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## Laboratory report

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APPLICATION LABORATORY-TAIPEI



For your personal information

number : NBA8105

date 17-11-1981

title : The MOS transistor BF982 in an  
FM preamplifier for car radio.

author : A. Hanck

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Electronic Components and Materials Division  
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## laboratory report

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NIJMEGEN - THE NETHERLANDS

REPORT No: NBA 8105

AUTHOR: A. Hanck

PROJECT No:

DATE: 17.11.1981

TITLE

The MOS transistor BF 982 in an FM preamplifier for car radio

ABSTRACT

For the r.f. part of car radios a supply voltage of only 8.5V is available.

Because also an automatic gain control is required, the BF 982 is very well suitable for this application.

Advice Patents Dept.

d.d: 21-01-1982

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GV

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BL

Decision MAMO

d.d: 12-02-1982

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DATE: 14 JAN. 1982 MAMO:

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REPORT No: NBA 8105

AUTHOR: A. Hanck

PROJECT No:

DATE: 17.11.1981

TITLEThe MOS transistor BF 982 in an FM preamplifier for car radioSUMMARY

The BF 982 has been developed for the application in r.f. stages for TV tuners with low supply voltage (12V).

Because of the low spread of the d.c. characteristics the transistor is also suitable for low voltage FM applications.

An FM preamplifier stage with gain control has been designed for car radio with an available supply voltage of 8.5V.

As expected, the circuit has a good d.c. stabilisation (low current spread), while the total product BF 982 can be applied (no need for transistor selections).

The circuit diagram is given in fig. 1S.

Some measured data:

gain	: > 20dB
noise figure	: < 4dB
agc range	: $\approx 60\text{dB}$ at $V_{\text{agc}} = 0.5\text{--}5.5\text{V}$
total current consumption:	7.5mA

Advice Patents Dept.

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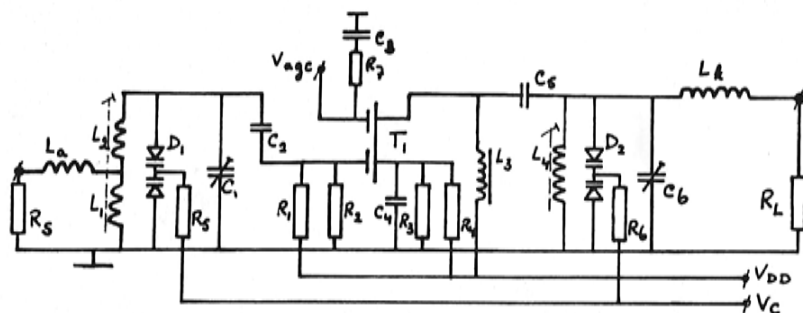
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$R_S = 75\Omega$

$C_1 = 2-10\text{pF}$

$L_A = 240\text{nH}$

$T_1 = \text{BF } 982$

$R_1 = 160\text{k}$

$C_2 = 2.7\text{pF}$

$L_1 \left. \begin{array}{l} + \\ - \end{array} \right\} \approx 102\text{nH}$

$D_1 = \text{BB } 204$

$R_2 = 100\text{k}$

$C_3 = 1\text{nF}$

$L_2 \left. \begin{array}{l} + \\ - \end{array} \right\} \text{tap } 1\frac{1}{2}:4\frac{1}{2}$

$D_2 = \text{BB } 204$

$R_3 = 470\Omega$

$C_4 = 1\text{nF}$

$L_3 = 5.5\mu\text{H choke}$

$V_{DD} = 8.5\text{V}$

$R_4 = 2\text{k}$

$C_5 = 100\text{pF}$

$L_4 = 111\text{nH}$

$V_C = 1.5-8\text{V}$

$R_5 = 56\text{k}$

$C_6 = 2-10\text{pF}$

$L_k = 1.2\mu\text{H}$

$V_{agc} = 0.5-5.5\text{V}$

$R_6 = 56\text{k}$

$R_7 = 33\Omega$

$R_L = 75\Omega$

fig 1S: test circuit diagram

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## 1. Introduction

Normally the publications and specifications of MOS transistors suggest the requirement of rather high supply voltages (15-20V). However, the BF 982 has been developed especially for amplifier stages with low supply voltages.

We have designed an amplifier stage for FM car radio front-ends with a supply voltage of 8.5V and automatic gain control.

## 2. DC adjustment of the amplifier stage

For the design of the d.c. circuit the following references are taken into account:

- the available supply voltage ( $V_{DD}$ ) is 8.5V
- the minimum agc voltage is 0.5V
- the drain voltage has to be 2V above the transistor knee voltage to have a sufficient excursion area.

From c) the maximum agc voltage can be calculated, see also fig. a.

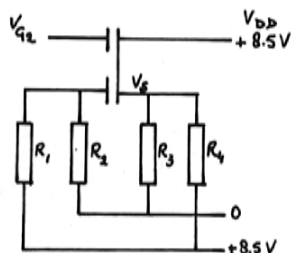


fig. a basic d.c. circuit

$$V_{kn} = V_{G2S} - V_{PG2} \quad (1)$$

$$V_{G2S} = V_{G2} - V_S \quad (2)$$

From (1) and (2) follows :

$$V_{G2} = V_{kn} + V_{PG2} + V_S \quad (3)$$

Also :

$$V_{DS} = V_{DD} - V_S = V_{kn} + 2V$$

$$\text{so: } V_{kn} = V_{DD} - V_S - 2V \quad (4)$$

From (3) and (4) follows :

$$V_{G2} = V_{DD} + V_{PG2} - 2V$$

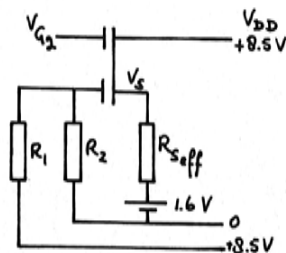


fig. b simplified d.c. circuit

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Because the limit of  $V_{PG2}$  of the BF 982 equals  $-1.1V$ , the nominal agc voltage has to be:

$$V_{G2} = 8.5 - 1.1 - 2 = 5.4V \quad (5)$$

At minimum agc voltage ( $+0.5V$ ) the MOST must be pinched-off ( $I_D = 0$ ). This means that the minimum voltage at the source ( $V_S$ ) has to be

$$V_{Smin} = V_{agcmin} - V_{PG2} = 0.5 + 1.1 = 1.6V \quad (6)$$

For an equivalent circuit, see fig. b.

In this figure the  $R_{Seff}$  stands for the parallel of  $R_3$  and  $R_4$ . The available voltage for d.c. stabilisation ( $V_{St}$ ) equals:

$$V_{St} = V_{G2} - V_{G2Snom} - V_{Smin} = 5.4 - V_{G2Snom} - 1.6 = 3.8 - V_{G2Snom}$$

For a good drain current stabilisation this  $V_{St}$  has to be as high as possible, which means that  $V_{G2S}$  has to be as low as possible.

Although in the specification of the BF 982 the  $V_{G2S}$  is always adjusted at  $4V$ , this is not a necessity in a practical circuit, see fig. 2. In this figure the transconductance versus the  $V_{G2S}$  is given with  $I_D$  as parameter. It may be concluded that for  $I_D = 10mA$ , the lowest value of  $V_{G2S}$  should be about  $2.8V$ . For  $I_D = 5mA$  resp.  $3mA$  the  $V_{G2S}$  has to be about  $1.9V$  resp.  $1.3V$ . At this choice of  $V_{G2S}$  also a certain spread of  $V_P$  is taken into account.

For these three cases we have calculated the d.c. adjustment:

- 1)  $I_D = 10mA$  and  $V_{G2S} = 2.8V$

$$V_{St} = 3.8 - 2.8 = 1.0V, \text{ so } R_{Seff} = V_{St}/I_D = 1V/10mA = 100\Omega$$

From fig. 1 follows that  $I_D$  can spread between  $6.2$  and  $14.3mA$

The source bleeder current can be calculated from:

$$R_{Seff} = \frac{R_3 + R_4}{R_3 + R_4} = 100\Omega \text{ while } \frac{R_3}{R_3 + R_4} \times 8.5 = 1.6V$$

This delivers:  $R_4 = 531\Omega$  and  $R_3 = 123\Omega$

$$I_{BL} = \frac{1.6}{R_3} = 13mA$$

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$$2) I_D = 5\text{mA and } V_{G2S} = 1.9\text{V}$$

$$V_{St} = 3.8 - 1.9 = 1.9\text{V} \rightarrow R_{Seff} = 380\Omega$$

$I_D$  spreads from 3.8mA to 6.2mA (see fig. 1)

The bleeder resistors:  $R_4 = 2019\Omega$  and  $R_3 = 468\Omega$

$$I_{BL} = \frac{1.6}{R_3} = 3.4\text{mA}$$

$$3) I_D = 3\text{mA and } V_{G2S} = 1.3\text{V}$$

$$V_{St} = 3.8 - 1.3 = 2.5\text{V} \rightarrow R_{Seff} = 833\Omega$$

$I_D$  spread from 2.4mA to 3.6mA (see fig. 1)

The bleeder resistors:  $R_4 = 4427\Omega$  and  $R_3 = 1027\Omega$

$$I_{BL} = \frac{1.6}{R_3} = 1.56\text{mA}$$

From the calculations above we have chosen for an  $I_D$  adjustment at 5mA. This delivers a good d.c. stability and a reasonably high transconductance.

The practical values for  $R_3$  and  $R_4$  will be 470 $\Omega$  resp. 2k $\Omega$ .

Now the gate bleeder can be calculated:

$$V_S = \frac{R_3}{R_3 + R_4} \cdot V_{DD} + \frac{R_3 + R_4}{R_3 + R_4} \cdot I_D = 3.52\text{V}$$

From fig. 1 follows that  $V_{G1S} = -0.22\text{V}$  at  $I_D = 5\text{mA}$ , thus the voltage at the gate has to be 3.3V.

This implies that  $R_2/R_1 = 3.3/(8.5 - 3.3) = 3.3/5.2$ .

We have chosen  $R_2 = 100\text{k}\Omega$  and  $R_1 = 160\text{k}\Omega$ .

It is also required that the d.c. circuit ensures a good agc curve.

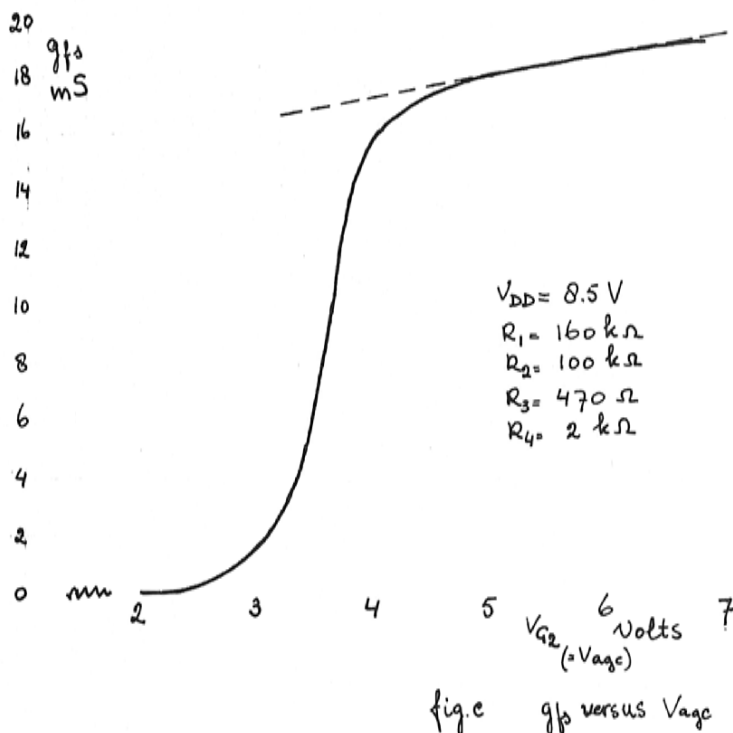
Therefore, we have measured the transconductance as a function of  $V_{G2}$  under the above calculated d.c. conditions, see fig. c. From this figure it may be concluded that the calculated d.c. circuit ensures a good agc curve.

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### 3. AC design of the test circuit

#### 3.1 The tuned circuits

In the tuned circuits the varicap BB 204 has been applied. The available control voltage is 1.5-8V. At these voltages the capacitance of the varicap varies from 23.3 to 12.4pF. To have some tuning reserve we have calculated the circuits for a frequency band of 86.2 up to 105MHz.

Calculated from  $\left(\frac{f_{max}}{f_{min}}\right)^2 = \frac{C_P + C_{Dmax}}{C_P + C_{Dmin}}$ , the parallel capacitance

$C_P$  has to be 10pF while an inductance of 102nH is required. Since the circuit damping is mainly determined by the series resistance of the inductance, the damping will be:

$$g = \frac{B_0}{2\pi f^2 L}$$

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while the unloaded bandwidth ( $B_0$ ) will be constant over the frequency band. To maintain a constant bandwidth the aerial and the load has to be inductively coupled to the tuned circuits.

### 3.2 Aerial circuit

The noise figure of the preamplifier is determined by the noise figure of the transistor and the available loss of the aerial tuned circuit. The available loss is given by

$$L_{av} = \frac{g_a + g_m}{g_a}, \text{ in which}$$

$g_a$  = aerial damping on the top of the circuit

$g_m$  = unloaded tuned circuit damping.

Because both dampings have the same frequency dependency, the available loss will be constant over the band. For a good large signal handling of the amplifier high selectivity of the tuned circuits will be required. This implies a rather high available loss and thus a poor noise figure. A low noise figure will be obtained if the available loss is very low, but then the aerial circuit will hardly deliver any contribution to the selectivity. As a compromise we have chosen for an aerial available loss of about 3dB. This means that the aerial damping equals the unloaded tuned circuit damping. It results also in a good voltage standing wave ratio (VSWR).

For practical reasons we have chosen a tuned circuit inductance with a tap (14:34) to avoid large values of the aerial coupling inductance. Experimentally the coupling inductance appears to be about 240nH. Applied inductances:

tuned circuit inductance: TOKO MC 111

coupling inductance : aircoil 9 turns  $C_u$  0.35mm  $\varnothing$  5mm

Measurements on the input circuit showed that the unloaded bandwidth was 0.93MHz, while the loaded bandwidth was 2MHz, and remains constant over the frequency band. This implies an available loss of the tuned circuit of 2.7dB.

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### 3.3 Drain tuned circuit

The drain is connected to the supply voltage via a choke and is capacitively coupled to the drain tuned circuit. For the same reasons as mentioned in chapter 3.1 also the load of the test circuit has been inductively coupled to the top of the tuned circuit. Measurements on the drain tuned circuit delivered:

The unloaded bandwidth of the tuned circuit is 0.81MHz.

The damping contribution of the choke is about 80μs and rather constant over the band (88-74μs at 86-105MHz).

We have chosen the load conductance equal to the tuned circuit damping (exclusive the choke damping). Now the coupling inductance can be calculated from:

$$g_L = \frac{R_L}{\omega^2 L^2 k} \text{ equals } g_x = \frac{B_0}{2\pi L f^2}, \text{ so: } \frac{R_L}{\omega^2 L^2 k} = \frac{B_0}{2\pi L f^2} \rightarrow L^2 k = \frac{R_L L}{2\pi B_0} \\ \rightarrow L_k = 1.2\mu\text{H}$$

A coil of 22 turns of enamelled Cu wire 0.35mm and a diameter of 5mm delivers about 1μH at low frequencies but due to its capacitance the active inductance appears to be about 1.2μH.

### 3.4 Coupling the transistor to the tuned circuits

Of a number of BF 982 the Y parameters at  $V_{DS} = 5\text{V}$ ,  $V_{G2S} = 2\text{V}$  and  $I_D = 5\text{mA}$  have been measured at 100MHz. The average values are given below:

$$Y_{is} = 0.1 + 2.4j\text{mS} \\ Y_{fs} = 18.2e^{-j10}\text{mS} \\ Y_{rs} = 0.018e^{-j90}\text{mS} \\ Y_{os} = 0.06 + 1.35j\text{mS}$$

In practice the printed board will deliver additional capacitances which normally may be ignored, except for the feedback capacitance, because the transistor  $C_{rs}$  is only 30fF. The extra capacitance amounts about 20fF, so it is better to do the stability calculations with  $C_{rs} = 50\text{fF}$ .

The circuit has the lowest stability factor at 105MHz, because there the tuned circuit dampings are minimum while the feedback admittance is maximum.

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The dampings are:

aerial damping (exclusive transistor) = 0.28mS

interstage damping (exclusive transistor) = 0.31mS

If the transistor is straight coupled to the tuned circuits the amplifier will be insufficiently stable, so it is necessary to tap the transistor to at least one of the tuned circuits. A tap on the aerial tuned circuit will decrease the gate input voltage and this has the advantage to increase the signal handling.

Another advantage of tapping is the decreased influence of the transistor parameters on the tuned circuits. These parameters will change during gain control, especially the input capacitance.

For both reasons we have chosen for a tap to the aerial circuit. As described in report NTI 8101<sup>x</sup> the gate is connected to the top of the aerial circuit by means of a small capacitance. A capacitance of 2.7pF will deliver a sufficiently stable amplifier at the cost of only a small increase of the noise figure.

The complete circuit is given in fig. 2.

Note that in series with  $G_2$  a resistor of  $33\Omega$  has been connected. This resistance will, instead of a ferrit bead, avoid parasitic oscillations.

#### 4. Measurements on the amplifier stage

The following measurements have been done:

transducer gain	fig. 4
noise	fig. 4
bandwidth (-3dB)	fig. 4
gain control	fig. 5
detuning during gain control	fig. 6
bandwidth during gain control	fig. 6

<sup>x</sup> NTI 8101: The stability factor of a selective amplifier with MOS transistor

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### 5. Conclusion

In the designed circuit the total product BF 982 can be applied, so a special transistor selection is not required.

The amplifier with BF 982 gives a low noise figure, high gain and a good gain control curve.

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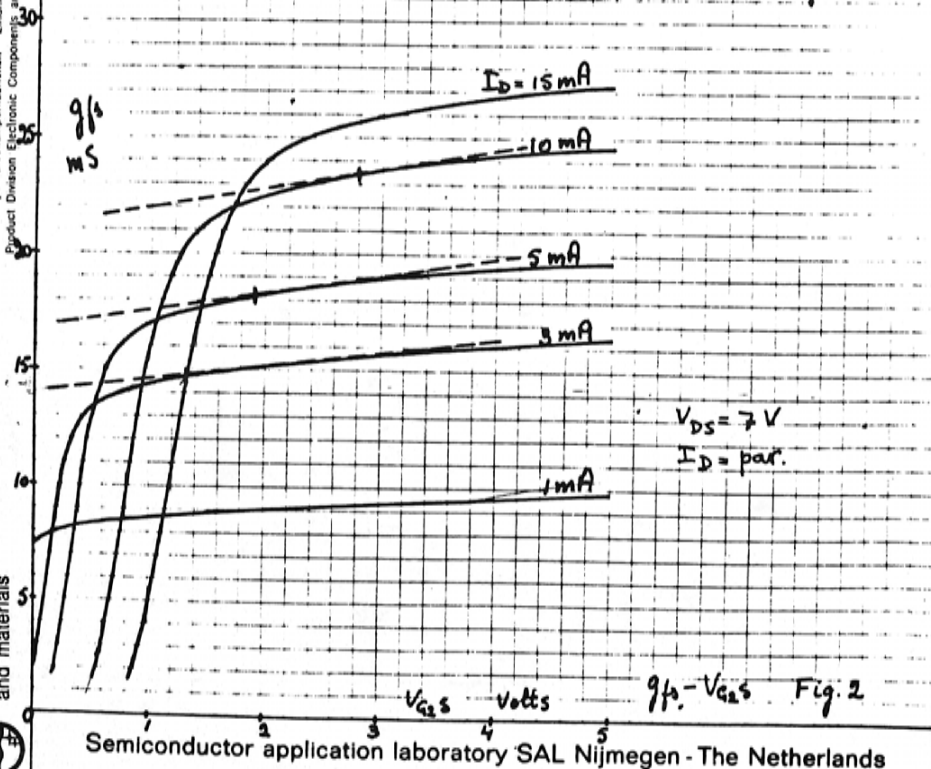
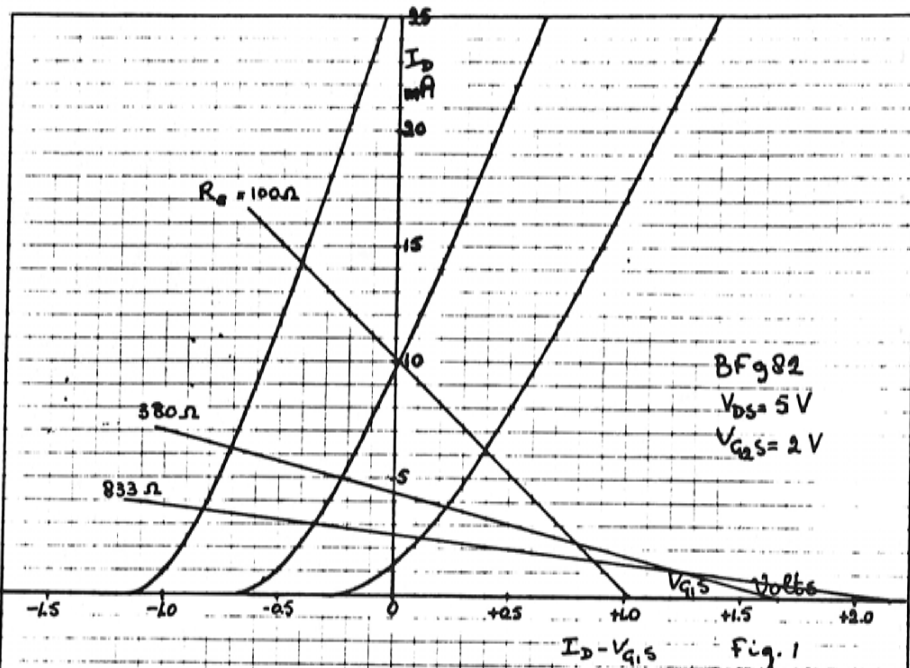
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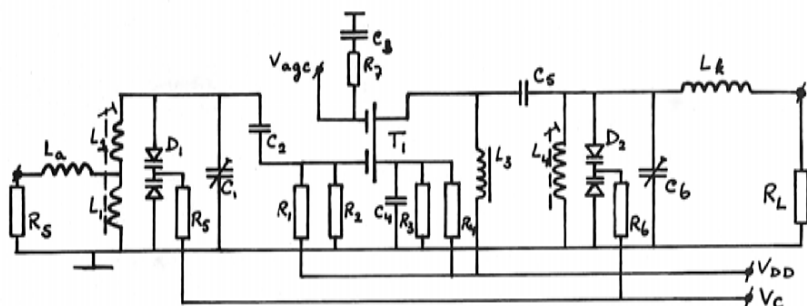
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$R_S = 75\Omega$

$C_1 = 2-10\text{pF}$

$L_A = 240\text{nH}$

$T_1 = \text{BF 982}$

$R_1 = 160\text{k}$

$C_2 = 2.7\text{pF}$

$L_1 \approx 102\text{nH}$

$D_1 = \text{BB 204}$

$R_2 = 100\text{k}$

$C_3 = 1\text{nF}$

$L_2 \left. \begin{array}{l} \uparrow \\ \downarrow \end{array} \right\} \text{tap } 1\frac{1}{2}:4\frac{1}{2}$

$D_2 = \text{BB 204}$

$R_3 = 470\Omega$

$C_4 = 1\text{nF}$

$L_3 = 5.5\mu\text{H choke}$

$V_{DD} = 8.5\text{V}$

$R_4 = 2\text{k}$

$C_5 = 100\text{pF}$

$L_4 = 11\text{nH}$

$V_C = 1.5-8\text{V}$

$R_5 = 56\text{k}$

$C_6 = 2-10\text{pF}$

$L_k = 1.2\mu\text{H}$

$V_{agc} = 0.5-5.5\text{V}$

$R_6 = 56\text{k}$

$R_7 = 33\Omega$

$R_L = 75\Omega$

fig. 3: test circuit diagram

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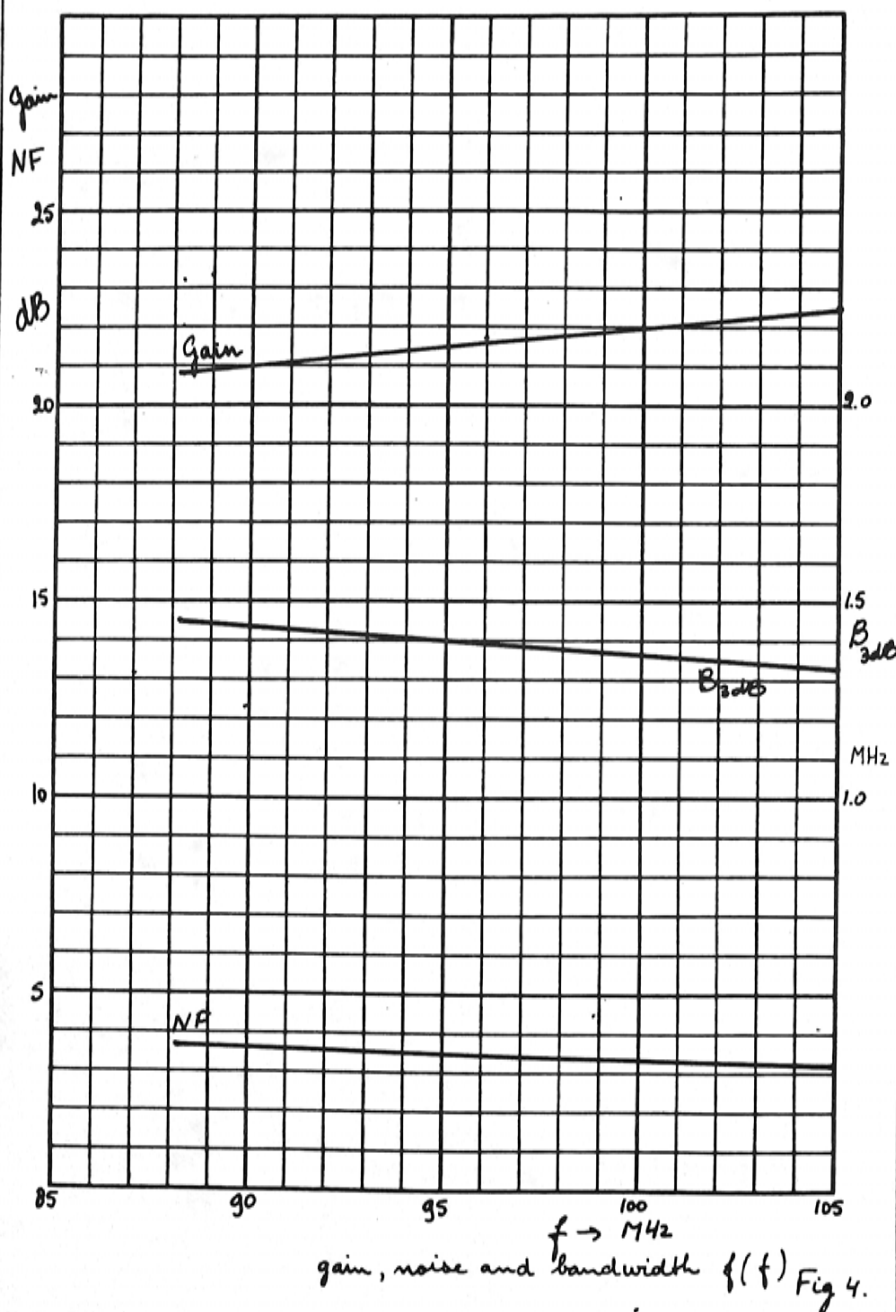
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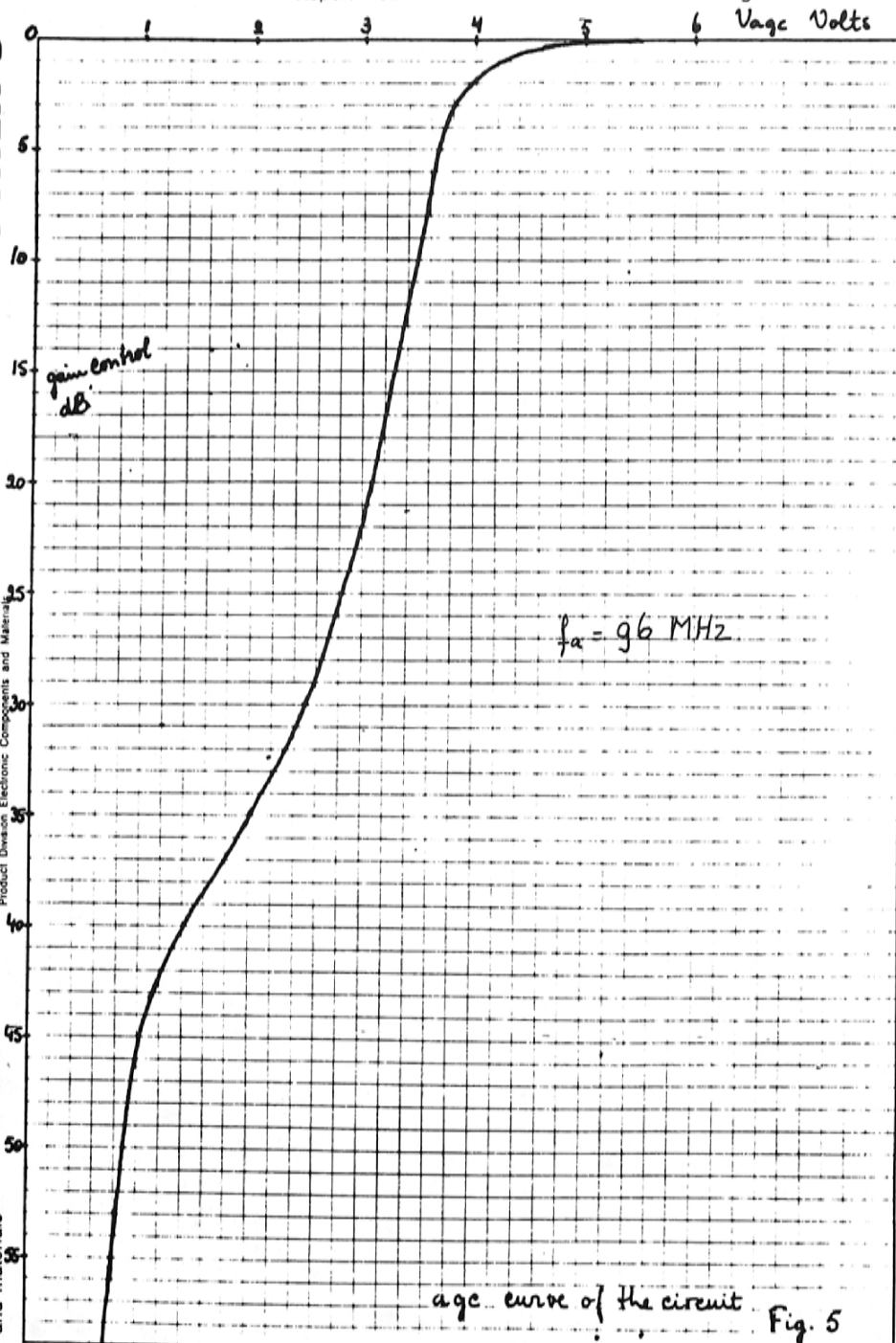
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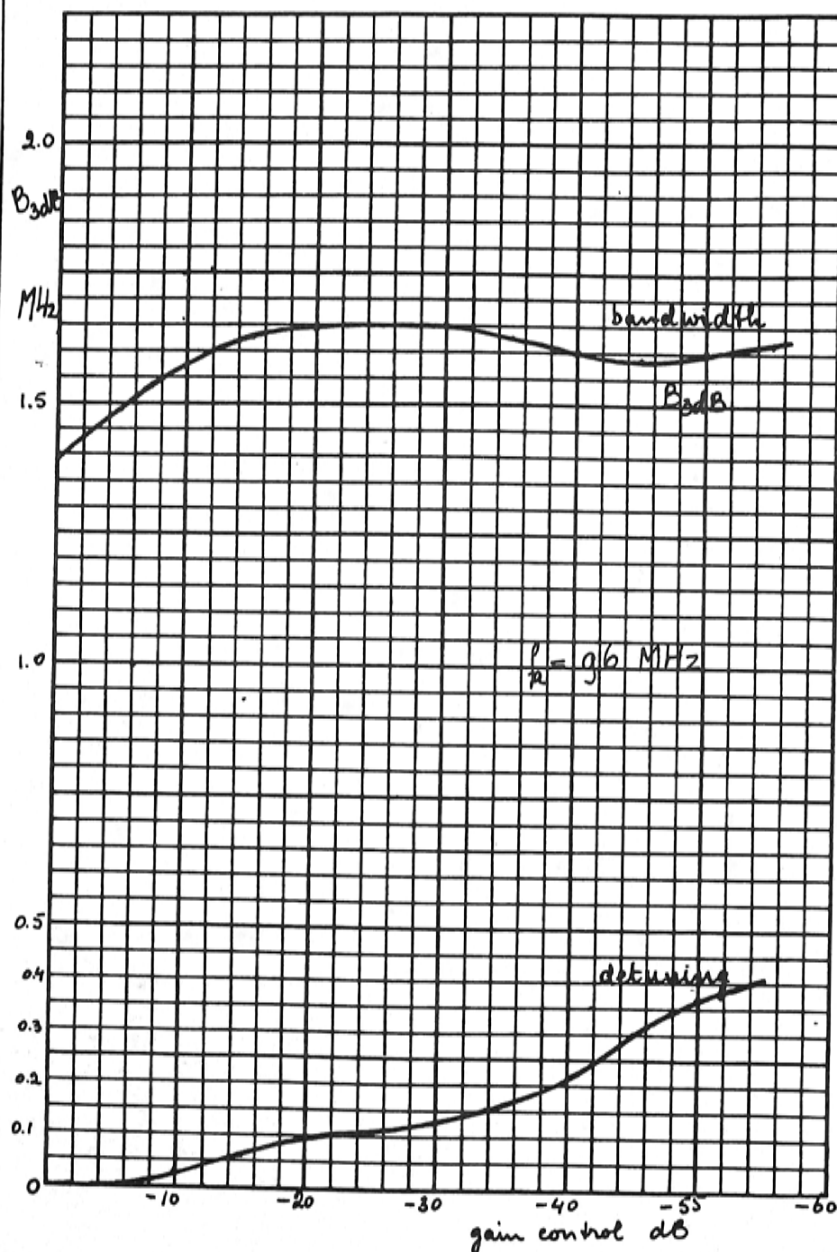


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Bandwidth and detuning  
versus gain control. Fig. 6

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